DEFENSE RESEARCH LABORATORY

THE UNIVERSITY OF TEXAS · AUSTIN 12, TEXAS

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SCATTERING CROSS SECTIONS OF MICROWAVE ANTENNAS

by

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THESIS

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Abstract

The effective radar cross sections of a number of horn and parabolic reflector microwave antennas were measured. X-band radiation was used throughout, but antennas designed for use at C-, X-, and K-band were measured. A low-power radar was used, designed for indoor operation and possessing a feature whereby the return from background objects was nulled out.

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I. INTRODUCTION

The broad technology which has been developed since the inception of radar has provided considerable information regarding the nature and character of microwave components. Without too much difficulty one might predict the performance of a section of hollow waveguide used as a transmission line, or perhaps, the radiation pattern of a parabolic reflector or a pyramidal horn used as a microwave antenna. There seems to be, however, a scarcity of published information concerning one facet of microwave antennas. This would be in regard to the radar of scattering cross section of such devices.

Because of the apparent lack of information on this subject, the prime objective of the study herein described was to experimentally establish the scattering cross sections of several parabolic and horn type antennas. Several antennas differing in size and configuration were illuminated with X-band radar energy. Some of the test antennas were designed for X-band, some for C-band, and some for K-band.

The complex geometry of most objects whose scattering cross sections might be of interest makes analytical evaluation difficult. As a consequence, empirical results are relied upon in most cases. Due to the comparatively simple geometry of the antennas selected for this study, however, the scattering cross sections will also be discussed from an analytical standpoint.

In addition to the principal objective of the study, the examination of antenna scattering cross sections, information of a different nature was also sought. The matter of concern here was the merit in the idea of performing antenna tests indoors without having to confine the test area to an anechoic chamber. A good portion of the program was thus devoted to the development of

a radar which might give plausible results under such conditions. A description of the equipment and the technique used to accomplish this is given early in the report and the merit of such a device is considered in the discussion of results.

II. IMPLEMENTATION

A. Test Antennas

In keeping with the stated objective of this study, the following antennas were constructed:

1. Two 18-inch aperture parabolic reflectors - one designed with an X-band feed and one with a C-band feed,
2. Two 12-inch aperture parabolic reflectors - one designed with an X-band feed and one with a K-band feed,
3. Three pyramidal horns of aperture 6 inches by 4 inches and axial length 11.75 inches - one for each of the C-, X-, and K-bands.

Several of these antennas are shown in Figure 1 along with samples of the waveguide components also used.

The measurement of the scattering cross section that each antenna presents in an operational situation is the parameter of interest in this paper. This does not require a complete receiving or transmitting system in the laboratory in order for the measured scattering cross section to be realistic. What is important is the terminating impedance of the antenna. Capitalizing on the fact that the energy intercepted by the target antenna was not to be utilized by any receiving system coupled to the antenna, the effect of a receiver was simulated by substituting an equivalent terminating impedance. Four different terminations were tried for each antenna in order that the effect on the scattering cross section might be examined for various degrees of matching between the antenna and the simulated receiver.

The antenna was deemed to be properly terminated when the termination was a piece of shorted waveguide in which was mounted an aquadag coated phenolic resin strip similar to bakelite. The strip, having a thickness of

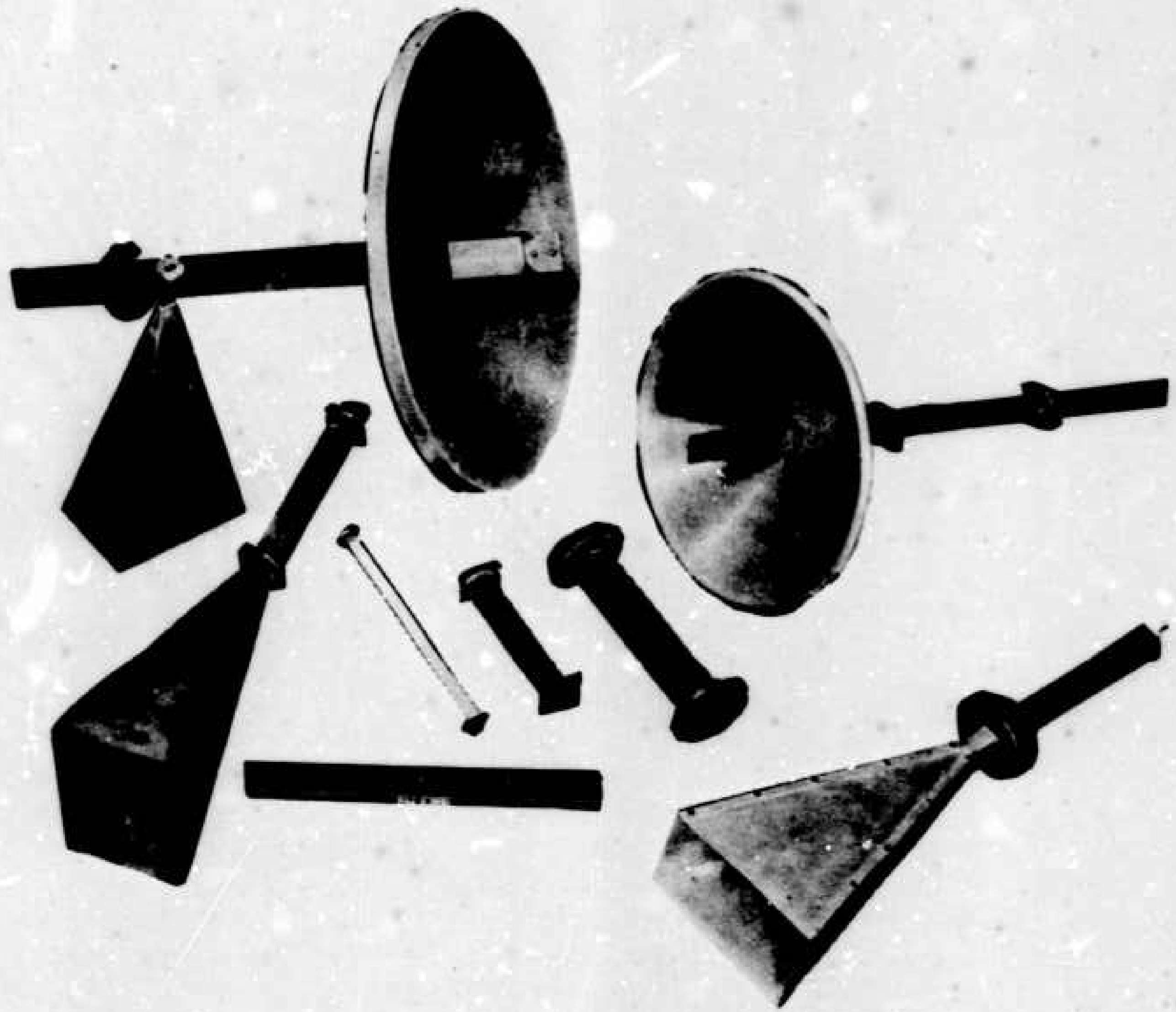

FIGURE I
TEST ANTENNAS

about 1/32-inch, was mounted parallel to the electric flux lines of the dominant TE_{10} waveguide mode and situated mid-way between the sides of the guide where the flux intensity for this mode is a maximum. The end of the absorbing strip nearest the antenna was tapered to minimize reflections. Waveguide devices using similar techniques are in common usage. As an indication of the ability of this device to match the waveguide, the voltage standing wave ratio was measured for a test line using this same termination and was found to be approximately 1.04 at 9375 Mc/s. The magnitude of the voltage reflection coefficient from this termination was thus approximately 0.02.

The terminating impedance providing the maximum impedance mismatch for the antenna was a shorted section of waveguide. The shorting section was constructed so that it could be mechanically adjusted to provide maximum reflection for the various waveguide lengths on the various antennas. Adjustable sections such as this were tested with the X-band and the C-band antennas.

Two additional loading arrangements were used to examine conditions of impedance between the extremes of the proper termination and the short circuit. A tuning probe was placed parallel to the electric field lines and located one quarter-wavelength from the closed end of the short circuited section. Direct waves traveling down the guide were reinforced by the reflected field at the probe position and the power coupled out of the guide was a maximum. This standard-type waveguide-to-coax coupler was left unloaded for one case and was connected to a 50 ohm resistor for the other. It was anticipated that the results using the unloaded probe would be similar to those of a shorted guide while those of the loaded probe would more nearly approximate the proper termination.

The only requirement on the length of the waveguide transmission line connecting the termination and the antenna itself was that it be long enough that higher order modes not normally propagated down the line would not influence the results. Since modes not normally propagated decay exponentially with distance down the guide, the desired condition could satisfactorily be met as long as the guide was more than a couple of wavelengths long.

A two-dipole feed of conventional design was employed for all the parabolic antennas. Since design information was available only for the X-band feed, it was necessary that the C- and K-band antenna feeds be scaled from this by a factor inversely proportional to the difference in center frequency of the desired antenna from the X-band center frequency.

With the double dipole feed, the driven dipole, the one nearest the waveguide, is situated at the focus of the parabolic reflector. The focal length of the 18-inch reflector was 5.4 inches while that of the 12-inch reflector was 4 inches.

B. Cross Section Reference

In addition to a comparison of the returns from the various antennas, the establishment of the absolute scattering cross section of each antenna was desirable. In order that this might be accomplished, a sphere was used as a reference target to which each antenna was compared. Any object might have been so used, but by using a sphere, difficulties involving orientation with respect to the transmitter-receiver antennas were reduced. In addition, the use of a sphere for a reference was quite compatible with the requirement for establishing radar cross sections, since the radar cross section has been defined as the "cross-sectional area of a perfectly conducting sphere

that would return the same power to the radar set as does the actual object."*
In this report this may be referred to as a radar cross section, scattering cross section, or just cross section.

According to the Friis Transmission formula or, more commonly, the radar range equatio.., the power reaching the receiver is a function of transmitter parameters, target range, target cross section, and the effective aperture of the receiving antenna. The received power might be expressed as:

$$P_r = \frac{P_t\ G_t}{(4\pi\ r^2)^2} \times A_r \times A_o \qquad (1)$$

P_t = transmitter power

G_t = transmitter antenna gain with respect to an isotropic antenna

r = range from transmitter-receiving to target

A_r = receiving antenna effective aperture

A_o = radar cross section of target

For the tests being discussed here, the range and the parameters of the transmitter-receiver unit (t-r unit) remained fixed. Variations in the received power were related solely to variations in the target cross section.

*Reintjes, J. F., and Coate, G. T., Principles of Radar, McGraw-Hill Book Co., Inc., N. Y., 1952, p. 20.

Taking a ratio for two separate targets then,

$$\frac{P_{r1}}{P_{r2}} = \frac{A_{o1}}{A_{o2}} = \frac{(V_{r1})^2}{(V_{r2})^2}$$

where V_{r1} and V_{r2} are the respective receiver voltages for for targets 1 and 2.

$$A_{o1} = A_{o2}\left(\frac{V_{r1}}{V_{r2}}\right)^2 \qquad (2)$$

If, then, the cross section of an object were known as were the receiver voltages which resulted from reflections from this object and from an object of unknown cross section, then the radar cross section of the unknown object might be established rather easily. This provided the basis for the selection of a sphere as the reference target.

Microwave energy incident upon a perfectly conducting sphere is reflected or scattered uniformly in all directions provided that the sphere diameter is large compared to the wavelength of the impinging energy. In such a situation, the sphere scattering cross section is equal to the projected area of the sphere (i.e., the area of a great circle of the sphere). The sphere employed, shown in Fig. 2, had a diameter of 16.96 inches or 43.1 cm. At the signal frequency of 9375 Mc/s, this diameter was roughly 13.5 wavelengths. The scattering cross section could thus be expected to be quite close to the projected sphere area.

$$A_{o2} = \frac{\pi d^2}{4}$$

$$A_{o2} = 1.57 \text{ square feet} \qquad (3)$$

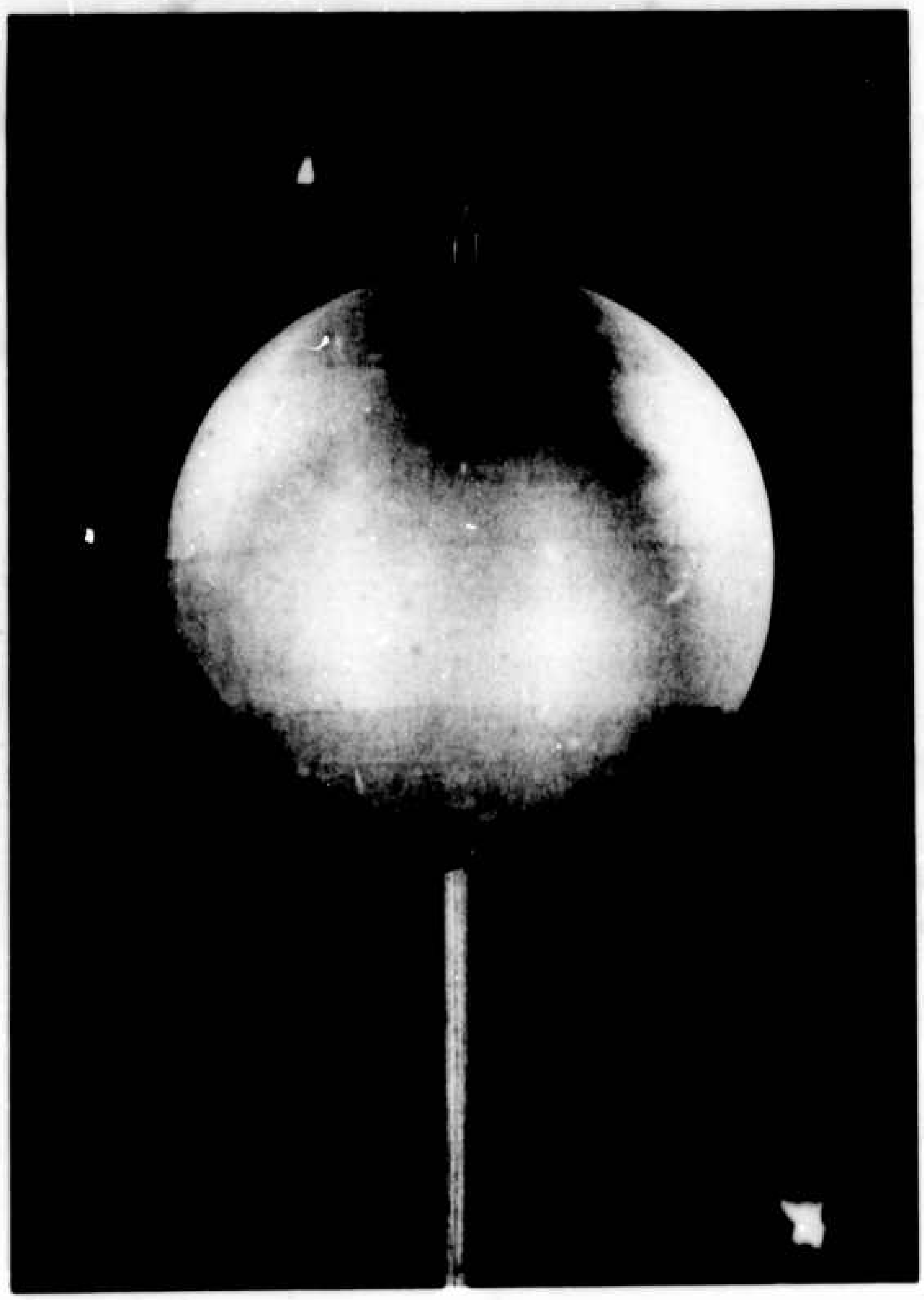

FIGURE 2
REFERENCE SPHERE

C. Indoor Measurement Technique

The measurement of antenna scattering cross sections in this study was performed on an indoor test range. Using a conventional radar system, the reflection of signals from the walls or from objects in the room might be expected to mask the signal of interest if the tests were performed in some enclosed area other than an anechoic chamber. Rather than build such a chamber to simulate free space test conditions, another technique was used and is discussed below.

A transmitter-receiver (t-r) unit was designed such that returns from fixed objects other than from the target antenna could be cancelled. The scheme used to cancel the undesirable returns from the background will now be described. As shown in Fig. 3, energy was supplied by a cw-transmitter to the transmitting antenna (the horn on the left in Fig. 3). By using a 3 db waveguide coupler, a portion of the energy was coupled out of the guide before it reached the antenna and was passed through an adjustable attenuator and an adjustable rotary phase shifter. This energy which was sapped from the transmitter signal was then coupled into the waveguide which also carried the returning signal picked up by the receiving antenna.

With proper adjustment of the attenuator and the phase shifter, it was found that the signal coupled from the transmitted signal could be made approximately equal in magnitude and 180 degrees out of phase with that signal returning through the receiver antenna. For this condition, therefore, the two signals would cancel. If an object whose scattering cross section was to be established were then placed in the antenna beam and the settings of the attenuator and phase shifter were not altered, then the energy entering through the receiving antenna would change and the previous condition

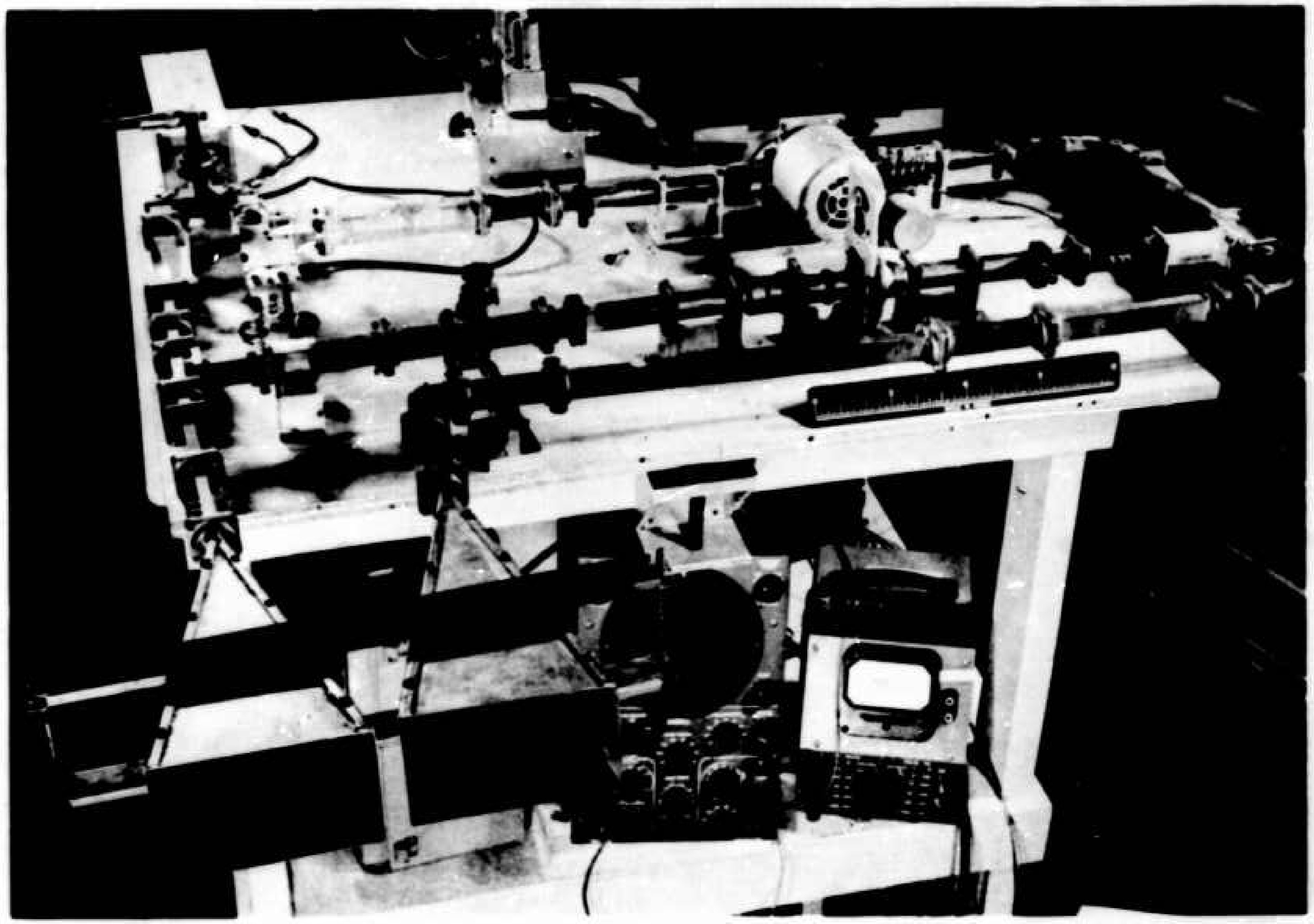

FIGURE 3

RADAR TRANSMITTER-RECEIVER UNIT

of balance would be upset. The signal increase, as a consequence of placing an object in the antenna beam, was then a measure of the cross section of that object.

The combined target and background signal was passed through a motor driven rotary phase shifter, which provided audio frequency phase modulation, and mixed with additional signal coupled directly from the transmitter signal. This was detected and the resulting audio signal amplified and presented on an oscilloscope and a voltmeter. Shown in the upper part of Fig. 3 are the preamplifier, the amplifier, and the power supply, moving from left to right. The transmitter is located on the shelf below the waveguide network.

It was stated that the system was adjusted for a null condition rather than a condition of zero output prior to measuring the target antenna. This was the result of noise inherent in the system. The system could, however, be adjusted until this noise level was about 3 db below the signal returned by the sphere. Stub tuners were used at several points to assist in minimizing this noise.

Since the tests were conducted indoors, limited use of microwave absorbent material similar to that used in anechoic chambers was deemed necessary. During testing, the antennas were mounted on a rotatable pole 3/4 inch in diameter and slightly less than six feet high. This is illustrated in Fig. 4 with a 12-inch diameter parabolic antenna mounted.

The procedure followed in taking data was to first mount on the antenna pole a piece of microwave absorbent material equal in area to the projected area of the antenna to be tested. The balancing attenuator and phase shifter

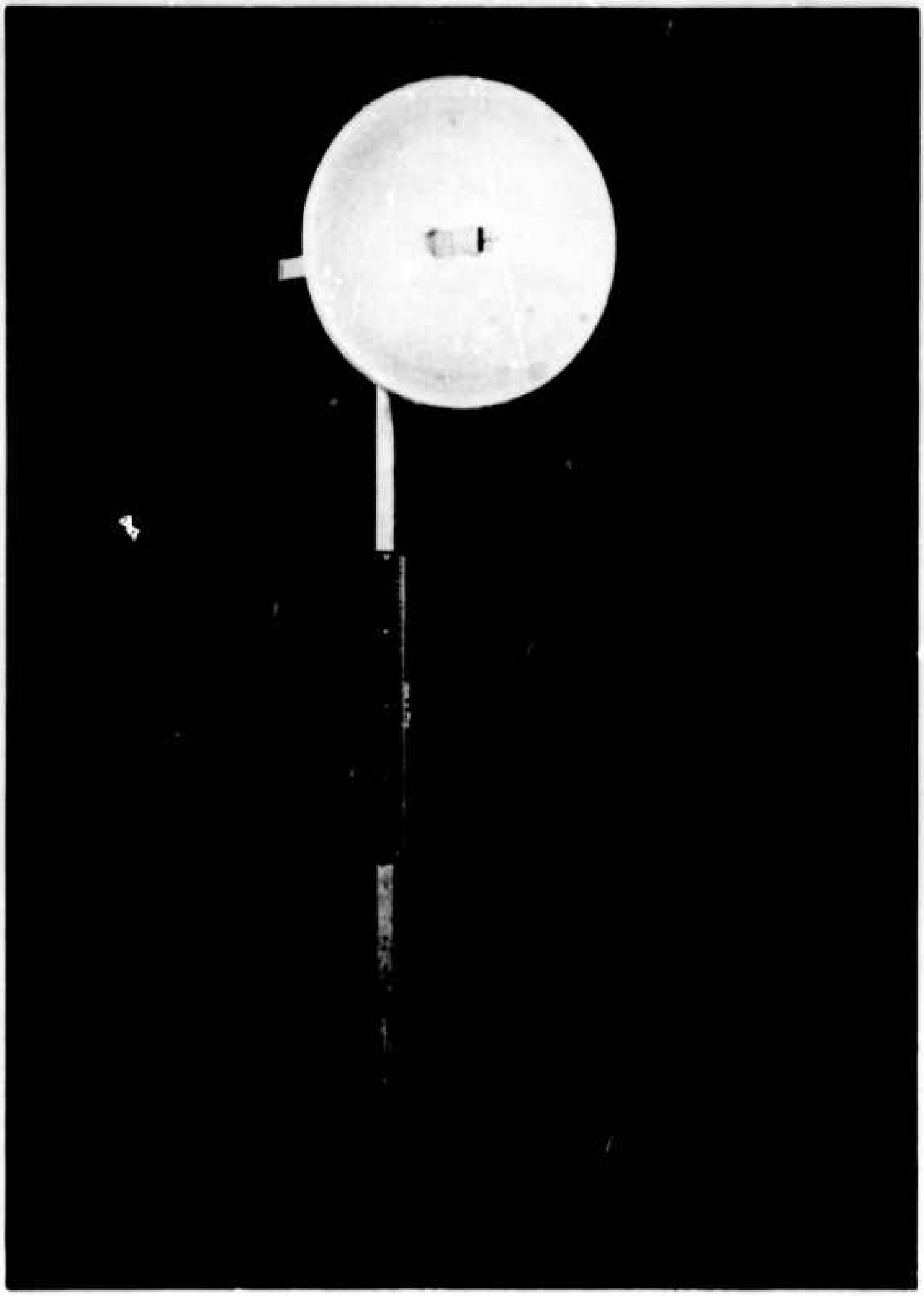

FIGURE 4

ANTENNA POLE MOUNT

were then adjusted to create a null in the output signal of the receiver. Following this the antenna was mounted and measured. The antenna was then removed and another piece of absorbent material, this time in the shape of a circle of a diameter equal to that of the reference sphere, was mounted and the radar system was again nulled. The reference sphere was then substituted for the absorber and the reflected signal was measured.

Had microwave absorber not been substituted for the antenna or sphere each time the system was nulled, then a slight error would have resulted from the contribution to the background signal by that region directly behind the antenna. The measured signal from the antenna would then have been made up of signal actually from the antenna plus that shift from the null voltage resulting from the masking of a portion of the background. Each different sized antenna or sphere would provide a different amount of masking. As it was used, some error still remained since the absorber was not perfect. This error is of little consequence, however, since the reflected power is something less than 2 per cent of that incident and not all of this is reflected in the direction of the receiver antenna.

Scattering cross sections were measured with the test antenna aimed toward the transmitter-receiver antennas. Thus much of the signal reflected from the test antenna was directed back toward the t-r unit. Part of this signal might strike an object and be redirected back toward the test antenna, thereby altering the measured cross section. The significance of this multiple bouncing of the signal was investigated and is considered in the discussion of results.

Signal from the transmitter not intercepted by the test antenna also might strike background objects and be reflected back to the receiver, altering the measured cross section. However, this signal was the same with

the test antenna mounted as it had been with the microwave absorber mounted on the pole. The radar set was designed to null out this signal component.

A small voltage remained after the system had been nulled. Thus, this voltage value had to be subtracted from the value measured for a test antenna to get the voltage which might correctly be related to the cross section.

Microwave absorbing material was placed on the floor between the test antenna mount and the t-r mount to see if reflections from the floor were of any consequence. The change in the measured cross sections with or without the absorber was insignificant.

D. Radar Components

Aside from the selection and arrangement of the waveguide components, the only element of the transmitter-receiver unit which actually required any design effort was the audio amplifier used in the receiver. The design details of this unit will be presented in this section. Other components used will be mentioned and described briefly.

An AN/UPM-10 radar test set served as the transmitter. This unit is capable of generating X-band signals over the frequency range from 8500 to 9600 Mc/s. The only frequency utilized was 9375 Mc/s. The signal produced was an unmodulated continuous wave (CW). Built into this unit are a power meter for monitoring the output and a cylindrical cavity-type absorption frequency meter. A 2K25 reflex klystron is used as the RF oscillator. Operation on three voltage modes of the klystron is possible, depending upon the reflector voltage. The nominal output power for these three modes ranges from about 15 to 25 milliwatts. The power actually getting to the transmitter antenna was closer to 5 milliwatts for the plumbing arrangement used in this test.

As described previously and pictured in Fig. 3, a portion of the signal from the transmitter was coupled from the waveguide just before it reached the transmitting antenna. This coupled energy was then passed through an adjustable attenuator and an adjustable rotary phase shifter.

The attenuator made use of a slab of dielectric material covered with an absorbent coating. This device was similar to the waveguide termination used, except that the position of the dielectric in the attenuator was adjustable from one edge of the guide to the center. This slab, being parallel to the electric flux lines of the dominant TE_{10} mode, could thus be moved from a region of minimum electric field intensity to that of a maximum. By means of this adjustment, then, the amplitude of the signal passed through this section and coupled into the receiver waveguide was made equal to that entering through the receiver antenna.

An adjustable rotary phase changer of the type described by Fox* was employed in order that the phase of the signal which passed through the adjustable attenuator might also have a desirable phase characteristic. More specifically, it was essential that the phase for this signal be 180 degrees different from that of the signal returning through the receiver antenna. Upon achieving this phase condition, therefore, the combination of the two signals would produce an output minimum.

Briefly, this phase shifter is composed of three differential phase shift sections in tandem. On either end is a section, fixed in position, which converts linearly polarized TE_{10} fields into the circularly polarized TE_{11} mode, or vice versa. Moving in the direction of propagation, the linearly polarized

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*Fox, A. G., "An Adjustable Waveguide Phase Changer", Proc. I.R.E., Vol. 35, 1947, pp. 1489-1498.

signal is converted to a circularly polarized signal with fields rotating in the clockwise direction. The center section, mounted on bearings and adjustable in position, lets the polarization remain circular but changes the direction of rotation to counterclockwise. The third section then converts the signal back into linear polarization. The actual phase shifting is accomplished by slabs of dielectric material arranged in a particular fashion in the different sections. There is very little loss in the dielectric and, as a consequence, the device is capable of producing a signal arbitrarily variable in phase and of the same amplitude as the input signal, for all practical purposes.

The term circular polarization refers not to the shape of the electric and magnetic flux lines, but rather to the manner in which this pattern changes with time. A linearly polarized wave is one whose field pattern does not change direction with time; the cross sectional field pattern of a circularly polarized wave rotates in the plane of the cross section as time varies.

A second rotary phase shifter was utilized in the waveguide circuit to modulate the received signal just before it was mixed with signal sampled directly from the transmitter signal. This phase shifter was driven by a 1/50 hp induction motor at a rate of 1220 rpm, each revolution producing a phase variation of 720 degrees. Thus the shifter was operating at a rate of 2440 cycles per minute or 40.7 cps. This operation being continuous, the phase of the received signal was thus continuously modulated at a 40.7 cps rate.

This signal was then combined with signal from the transmitter in a crystal mixer. The 40.7 cps component of the signal, detected by a 1N23D crystal, was then fed to a transistorized preamplifier.

Before discussing the amplifiers used, it is noteworthy that undirectional transmission devices (Unilines) were used at a couple of points in the plumbing circuit, shown in Fig. 3. The use of these ferrite devices restricted the flow of RF signal to the proper channels and caused some improvement in the system noise level.

The signal from the crystal detector was fed to a transistor preamplifier, the schematic of which is shown in Fig. 5. This particular amplifier had been designed previously at DRL by L. L. Huggins.* The amplitude-frequency characteristic of this amplifier is shown in Fig. 6. The mid-frequency gain of this amplifier is about 106 db. In the band of interest for this application, the gain is down about 3 or 4 db from this, making it around 102 db.

Because of this high gain available from the preamplifier, the rejection of 60 cps interference as well as harmonics of the 40 cps audio signal found to be present could be given prime importance in the selection of the amplifier characteristic. Here again a proven circuit was available at DRL for use as the basic amplifier. The response of this amplifier, shown schematically in Fig. 8, was then modified through the use of series and feedback R-C circuits to achieve the necessary amplitude-frequency characteristic.

In block diagram form, the amplifier circuit finally settled upon would appear as in Fig. 7a.

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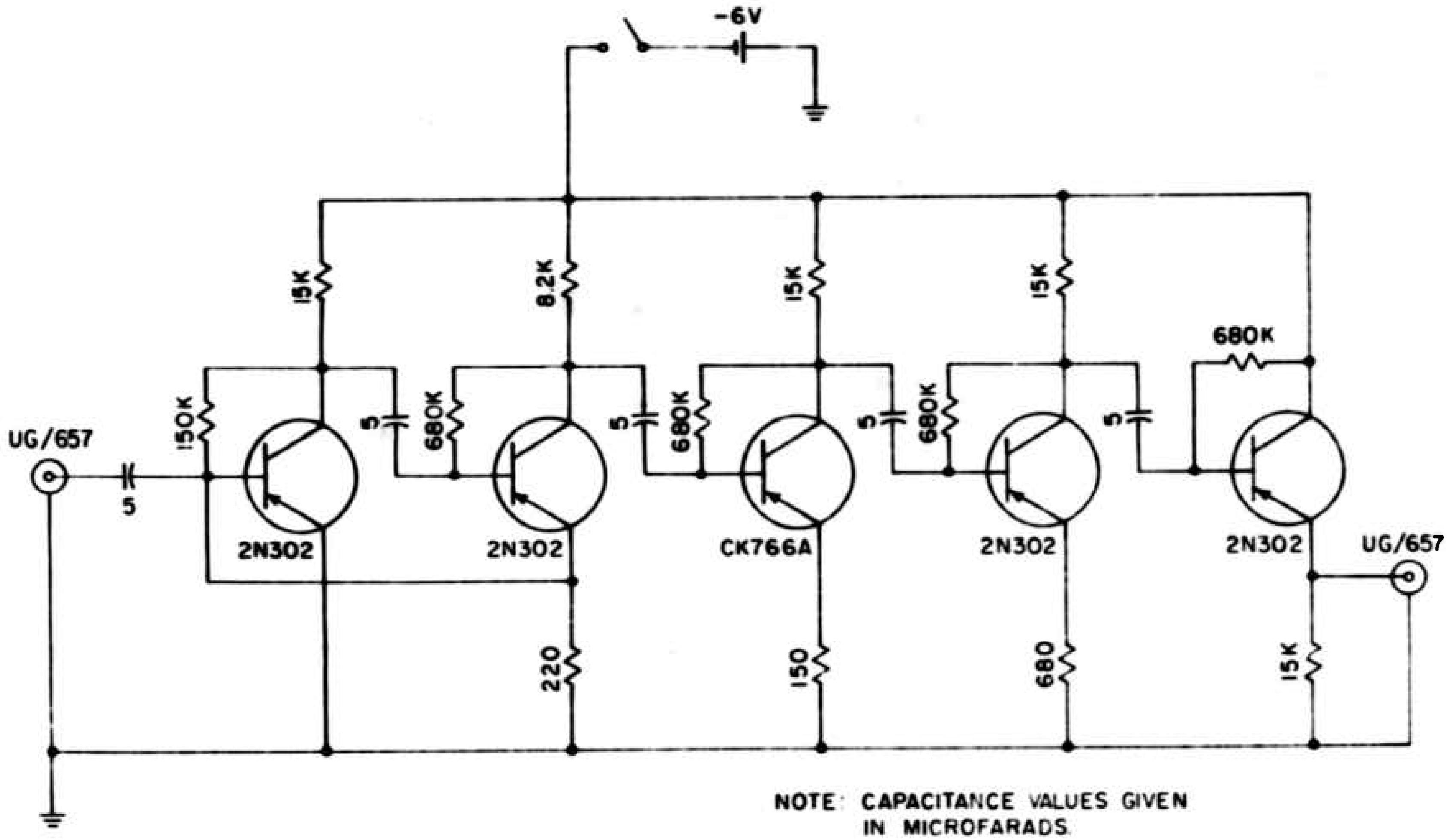
*Huggins, L. L., "A Practical System for Measuring FM Noise in Klystrons", Ph.D. Dissertation, University of Texas, 1959, pp. 20-24.



NOTE: CAPACITANCE VALUES GIVEN IN MICROFARADS

FIGURE 5
TRANSISTORIZED PREAMPLIFIER

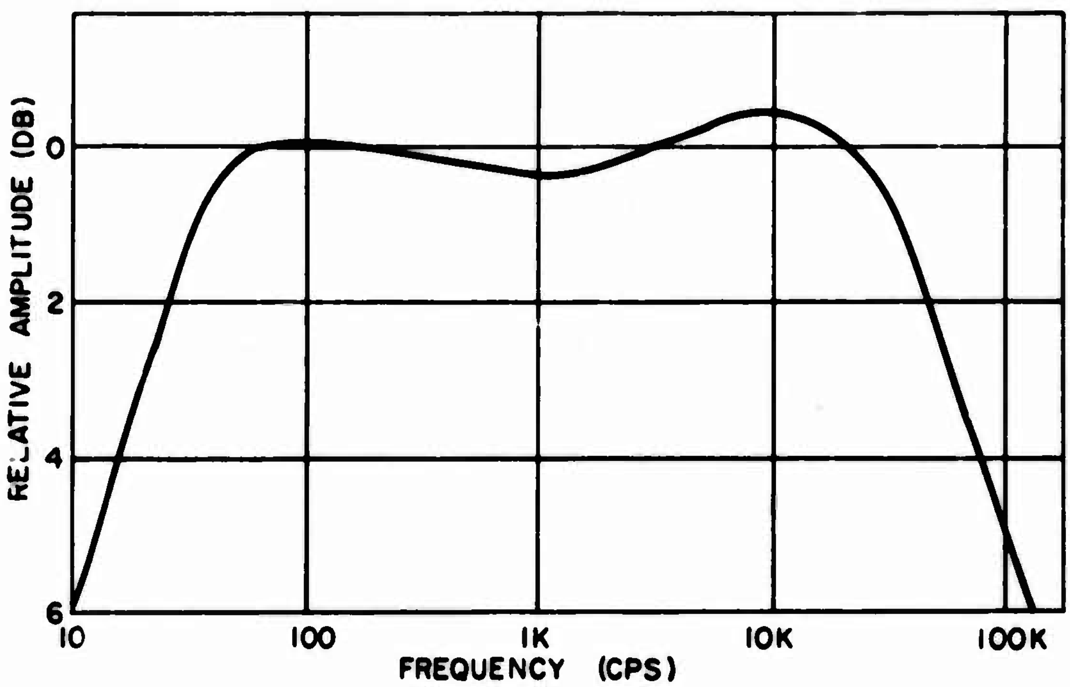
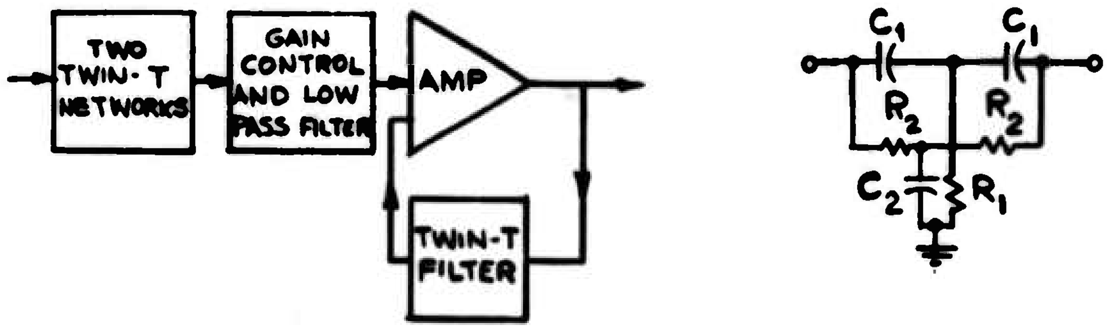


FIGURE 6

PREAMPLIFIER AMPLITUDE-FREQUENCY RESPONSE



(a) Amplifier Block Diagram (b) Twin-T Filter

Figure 7

The signal coming from the preamplifier was first passed through a twin-T circuit designed to provide rejection at 78 cps. In designing a twin-T circuit to provide rejection at some frequency, the following two equations must be satisfied.

$$2\, R_1 R_2 = \left(\frac{1}{\omega C_1}\right)^2 \tag{4}$$

$$R_2^{\,2} = \frac{2}{\omega^2\, C_1 C_2} \tag{5}$$

By choosing a relationship between R_1 and R_2 and a value for one of them, then the value of all components may be established at some design frequency. The elements of the twin-T are shown in Fig 7b. For rejection of the 78 cps signal the following values were used:

$R_1 = 100$ K

$R_2 = 200$ K

$C_1 = 0.01$ µf

$C_2 = 0.02$ µf

The condensers were slightly different from the design values and as a consequence the notch occurred at 77 cps. This change, of course, had no noticeable effect upon the end result.

The output of this series twin-T went to a second series twin-T designed for a notch at 160 cps. This filter was added to the circuit after noise at this frequency was experienced. For this filter,

$$R_1 = 100\ K$$
$$R_2 = 200\ K$$
$$C_1 = 0.005\ \mu f$$
$$C_2 = 0.01\ \mu f$$

The output of this second twin-T filter then went to a gain controllable potentiometer and then on to a low pass R-C filter designed for 60 cycles. The condenser available which was nearest in size to the design value changed this frequency to 72 cps. This filter prevented the amplitude of the output from increasing at frequencies above the twin-T series notches.

$C = 0.022\ \mu f$

$R = 100\ K$

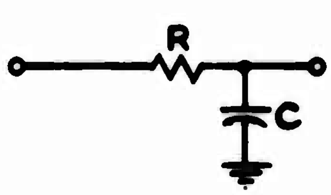


Fig. 9 Low Pass Filter

From this R-C filter, then, the signal went to the basic amplifier shown in Fig. 8.

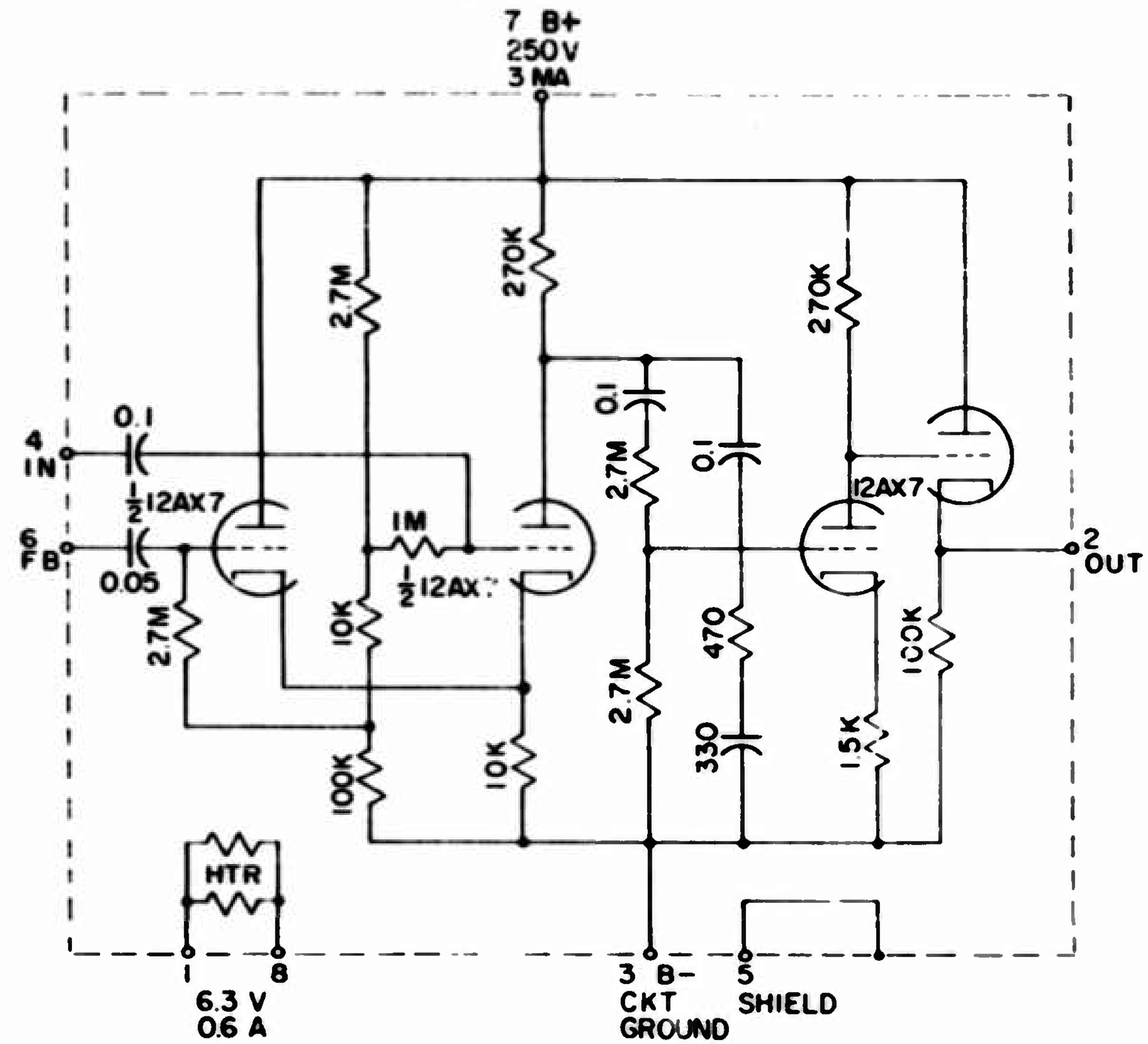


FIGURE 8
BASIC AUDIO AMPLIFIER

A feedback network had been used previously with this basic amplifier. Since narrow band response was now desirable, this negative feedback network was changed so as to contain another twin-T network, this one designed for 40 cps. The feedback network was connected between terminal 2, the output of the basic amplifier, and terminal 6, of Fig. 8. The network was designed to feed back all frequencies except 40 cycles. The diagram for this circuit is shown in Fig. 10.

For this feedback circuit,

R_1 = 47 K	C_1 = 0.02 μf
R_2 = 200 K	C_2 = 0.04 μf
R_3 = 100 K (Pot.)	C_3 = 700 μμf
R_4 = 1 M	C_4 = 120 μμf
R_5 = 85 K	
R_6 = 100 K	
R_7 = 2.6 K	
R_8 = 43 K	

The condenser C_4 was needed to prevent high frequency oscillations. The variable resistor in the twin-T arm made the network tuneable over a small frequency range. The amplitude-frequency response of the resulting audio amplifier is shown in Fig. 11, where amplitude is given in terms of db relative to the peak response at 30 cps. Beyond about 47 cps, the response drops off at a rate of approximately 82 db/octave. The resulting bandwidth was 30 cycles.

The output from this amplifier was then presented on an oscilloscope and a voltmeter.

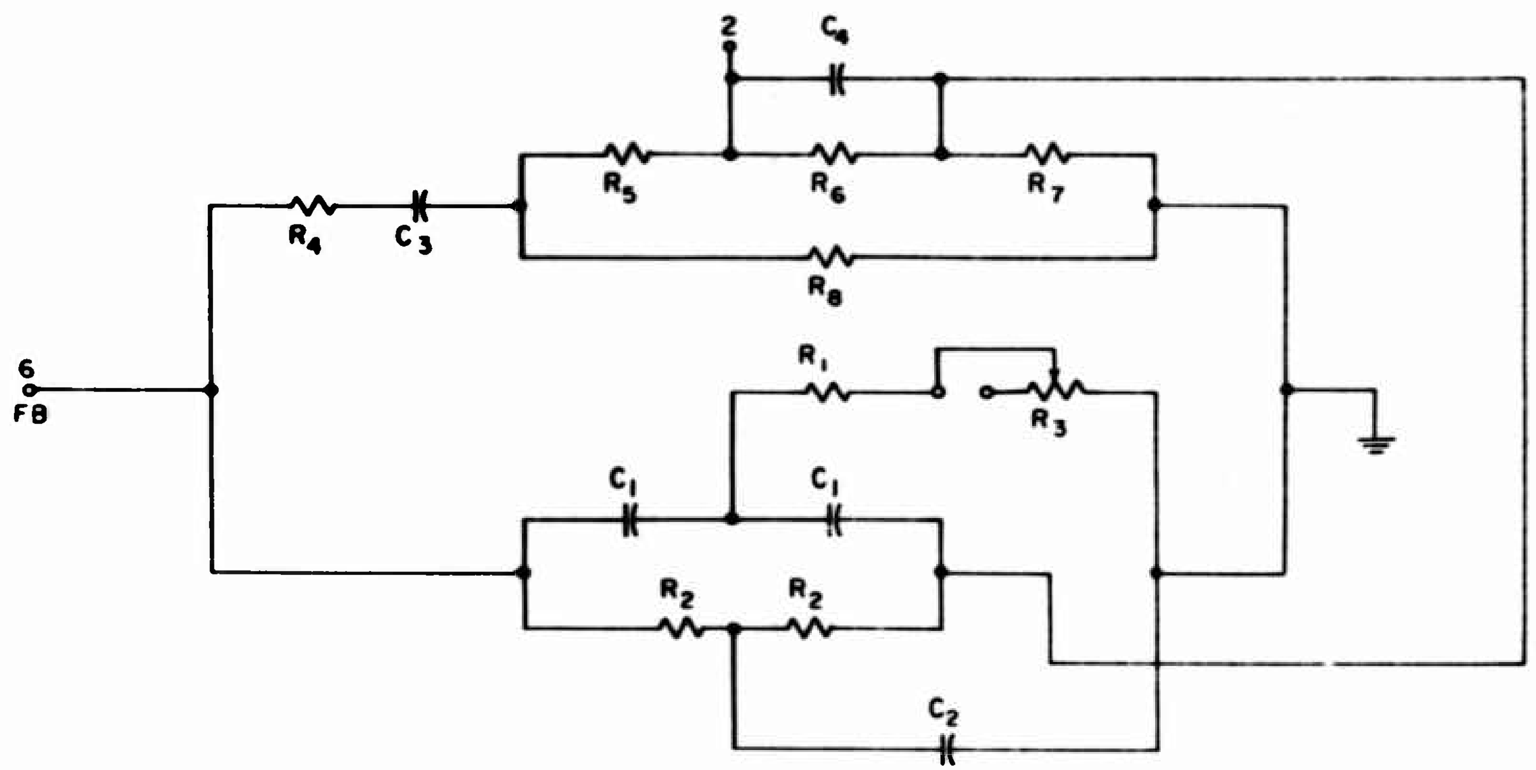


FIGURE 10
AMPLIFIER FEEDBACK NETWORK

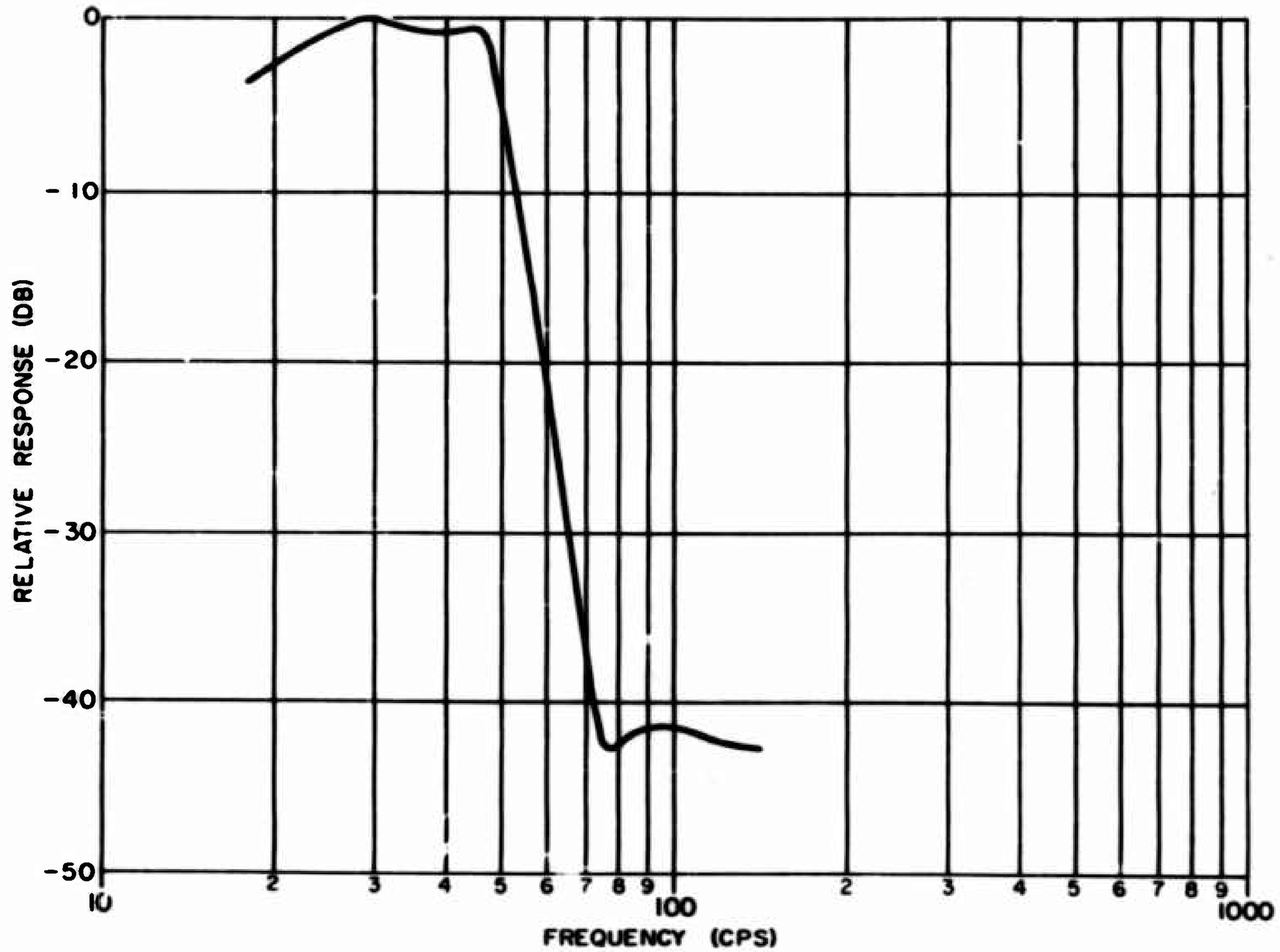


FIGURE 11

RESPONSE OF COMPLETE AUDIO AMPLIFIER

III. EMPIRICAL DATA

Using the procedure outlined in the preceding section, scattering cross sections were measured for the various antennas. The results are presented in Table I.

The scattering cross section values given in Table I were established with the target antenna centered or boresighted on a point midway between the transmitting and receiving antennas. Pointing the antenna a couple of degrees from the boresight position produced a significant change in the cross section.

The data for the various antennas are categorized in Table I according to the polarization of the target antenna relative to that of the t-r unit antennas. This polarization is called "coplanar" when in the plane of the horizontally polarized t-r horns and "cross" when in a plane normal to this. Each of these categories is then sub-divided into columns for the various antenna terminations. The scattering cross section of each antenna with each termination is given both in db above (+) or below (-) the cross section of the reference sphere and in terms of square feet (using equation 2).

For each termination and each antenna in the Table, an "N" or a "P" is indicated along with the scattering cross section. These indicate whether the antenna beam was at a minimum (null) or maximum (peak), respectively, with the antenna boresighted or centered on the t-r horns. With the antenna so oriented, beams of all antennas tested were found to be at one of the two positions - either a peak or a null. The existence of these peak and null values at the boresight position will be considered in the discussion of results. Also discussed in that section will be some of the factors which influenced the cross sections of the different antennas.

TABLE I. EMPIRICAL RESULTS

Antenna	Coplanar Polarized: Proper Termination		Coplanar Polarized: Probe (50 ohm load)		Coplanar Polarized: Probe (Unloaded)		Coplanar Polarized: Shorted Termination		Cross Polarized: Proper Termination		Cross Polarized: Shorted Termination	
	db	sq ft	db	sq ft	db	sq ft	db	sq ft	db	sq ft	db	sq ft
18" Parabolic (X-band)	N -1[illegible].6	.04	N -3.5	.7	P 33.2	3240	P 35.0	4965	P 21.0	199	P 21.0	199
18" Parabolic (C-band)	P 29.6	1436	P 29.8	1493	P 30.9	1922	P 31.4	2160	P 25.5	355	P 25.5	355
18" Dish (No feed)	-		-		-		P 18.4	109	-		-	
12" Parabolic (X-band)	P 25.8	381	P 24.9	485	P 28.3	1072	P 32.5	2770	N 14.0	[illegible]	N 16.1	65
12" Parabolic (K-band)	-		-		-		N 9.1	13	-		P 2[illegible].[illegible]	1[illegible]
Horn (X-band)	N 5.2	3	N 2.1	3	P 10.3	17	P 18.4	107	N 8.1	10	N 7.[illegible]	[illegible]
Horn (C-band)	N -2[illegible].8	.01	N -.8	1	N 5.6	6	P 20.1	1[illegible]0	P 2[illegible].8	300	P 23.[illegible]	[illegible]16
Horn (K-band)	-		-		-		P 19.9	15[illegible]	-		P 7.1	8

An attempt was made to measure antenna scattering cross section patterns by pointing the antenna at different angles relative to the transmitter-receiver unit, but little confidence could be placed in the results. First, since separate antennas were used for transmitting and receiving and at a range of twenty feet they could not be considered a point source, then a symmetrical target antenna would not be illuminated exactly the same when pointed to one side of the center position and then pointed to the other. Second, when the antenna was rotated energy was reflected from the antenna to objects in the background, back to the antenna, and back again to receiver. A variation in antenna orientation was, in effect, a variation in the background. Thus the actual change in the antenna scattering cross section was unintelligible. This was verified by mounting the antenna on the test pole, shown in Fig. 4, and measuring the pattern over an angular range of about 15 or 20 degrees to either side of the center or boresight position. The antenna was then rolled 180 degrees about the principal axis of the waveguide and the measurement repeated. Any dissymmetries in the antenna scattering cross section pattern might then have been expected to be reversed. For all cases tested, the pattern was always altered but not reversed exactly. It was thus presumed that the measurements could not be relied upon to give an accurate indication of antenna scattering cross section at off-center angles. The question of the validity of the cross section results with the antenna boresighted will be considered in the section on the discussion of results.

IV. DISCUSSION OF RESULTS

Looking again at Table I, the measured scattering cross sections are found to vary considerably for different antenna orientations and terminations. Since the measured values at off-center positions were of questionable accuracy, some assurance of the validity of the cross sections measured with the test antennas boresighted should be established before progressing further.

A. Ringing

One factor which might distort the measured scattering cross sections would be multiple bouncing of energy between the target antenna and the t-r unit. Rewriting equation (1) expressing the power absorbed by the receiving antenna due to energy which has gone from the transmitter to the target and back,

$$P_r = \frac{P_t \, G_t \, A_r \, A_o}{(4\pi \, r^2)^2} \, .$$

Of the energy impinging on the mount and the horns of the t-r unit, a portion is again returned toward the target. The receiver horn itself actually scatters or reflects more energy than it absorbs.* Thus the horns combine with the other structure of the t-r unit and mount to make up an object of scattering cross section A_{om}. The power density returned from the target must be multiplied by this cross section (A_{om}) to get what might be called the secondary-transmitter signal. This and subsequent bounces would combine with the initial power returned to provide an infinite series of terms which converge to the value of the power density at the receiver. For the multiple reflection or ringing effect to be negligible, all terms after the

*Kraus, J. D., Antennas, McGraw-Hill Book Company, Inc., N. Y., 1950, p. 47

first of this series must be negligible. Since these terms become progressively smaller, only the second term need be considered. For the combined transmitter signal and secondary-transmitter signal, the power absorbed by the receiving antenna now becomes:

$$P_r' = \frac{P_t G_t A_r A_o}{(4\pi r^2)^2} + \frac{P_t G_t A_o}{(4\pi r^2)^2} \times \frac{A_{om}}{4\pi r^2} \times \frac{A_o}{4\pi r^2} \times A_r$$

$$P_r' = \frac{P_t G_t A_r A_o}{(4\pi r^2)^2} \left[1 + \frac{A_o A_{om}}{(4\pi r^2)^2}\right] \qquad (6)$$

The scattering cross sections of both the target and the t-r mount are seen to influence the significance of the ringing effect. The separation of the two for these tests was 20 feet.

With the reference sphere as a target (A_o = 1.57 square feet), let the effect of ringing be examined.

$$P_r' = P_r \left[1 + \frac{1.57 A_{om}}{(4\pi)^2 (20)^4}\right]$$

$$P_r' = P_r \ (1 + 6.21 \times 10^{-8} A_{om})$$

If the scattering cross section of the mount were 10^6 square feet, the power absorbed by the receiving antenna would even then be increased by only about 6 per cent. Since a cross section of 10^6 square feet is only a factor of ten below that which has been estimated for large ships,* that ringing has a negligible effect for this case seems a valid assumption.

*Pollard, E. C. and Sturtevant, J. M., Microwaves and Radar Electronics, John Wiley and Sons, N. Y., 1948, p. 342.

The larger the scattering cross section of the target, the smaller the mount cross section has to be to produce a measurable effect on the result. As another example, take $A_o = 3000$ square feet.

$$P_r' = P_r (1 + 1.19 \times 10^{-4} A_{om})$$

Now a cross section, A_{om}, of only 1000 square feet could produce an error of about 12 per cent. This fact would seem to create some doubt as to the correctness of the data given for targets of large scattering cross section. It was thus essential that the order of magnitude, at least, of A_{om} be established.

The t-r unit contains two X-band pyramidal horns of the same size as the X-band horn listed in Table I. These horns were tuned to achieve the optimum impedance match between the waveguide and free space. The terminations for these horns should approximate the condition, also listed in the table, of a probe connected to a 50 ohm load. Since the measured cross section of this type of antenna so terminated was less than 10 square feet, it has been shown that there is reason to believe that this is a credible value, from a ringing standpoint at least. For the purpose of estimating the magnitude of A_{om}, therefore, it would seem that a reasonable safety factor had been provided if each horn of the t-r unit was assumed to have a cross section no greater than 50 square feet, for energy of either polarization. The influence of the remainder of the t-r mount can not be so easily estimated, since the structure is complex and difficult to describe analytically.

The technique employed for evaluating the effect of the mount was to again utilize microwave absorber material of the type described in the previous section of this report. First, an antenna was mounted and the cross section measured as before. Leaving everything else unchanged, then, absorber was

affixed to the t-r mount such that all components were masked except the horns themselves. Again making the assumption of negligible reflection from the absorber, the reduction in the measured target cross section was a measure of the t-r mount scattering cross section. As anticipated, the percentage reduction when the absorber was added was greatest for large antenna cross sections. The measured variations ranged from zero to about 6 per cent of that cross section with no absorber on the mount. For the shorted, 18-inch diameter, X-band, parabolic reflector, this change in A_{om} caused a change in the measured cross section of about 300 square feet. Additional error in the measured cross sections will be caused by the horns, which were not masked here. Assuming a cross section for the two horns of 100 square feet, the error would be increased by around 2 per cent.

Ringing is thus seen to cause a detectable change in the scattering cross sections measured for antennas of large scattering cross section, at a target range of 20 feet. The error should be less than 8 per cent of the corresponding value given in Table I. Such an error in a test of this type, however, does not render the data useless.

B. Radiation Patterns

As stated earlier, the measured antenna scattering cross sections at off-center angles were of doubtful validity. A few statements regarding the radiation pattern of the antenna used as a scatterer can be made, however, with some confidence.

First of all, the radiation pattern contains many more fluctuations than would normally be found in the free space pattern of the same antenna. Rather than a uniform main lobe devoid of inflections as might be anticipated from a horn antenna, for example, the radiation patterns of the antennas

tested were found to have a scalloped appearance. Used as a transmitter, the X-band horn tested, having an aperture of approximately 5 wavelengths, would be expected to have a beam width between first nulls of about 30 or 35 degrees. As a scattering target, this width was found to be on the order of 10 degrees. Several lobes were found to occupy the same angular space formerly occupied by the main lobe. The beam width between the first nulls for the parabolic reflectors was found to be about 5 or 6 degrees.

This debasement of the antenna beam pattern is due to the fact that all of the energy which is incident upon the antenna is not successfully coupled away from the antenna through the associated transmission line. A portion is scattered by such discontinuities as feed structures or edges of the collecting aperture. Also, for a parabolic reflector some of the energy focused onto the feed is returned due to the fact that the feed is not one hundred per cent successful in coupling energy into the transmission line and due to imperfect focusing in the primary reflector. A part of the energy successfully coupled into the transmission line is returned to the antenna and re-radiated because of a mismatch in the terminating impedance. The energy from these various sources is scattered with a given directivity (the ratio of the maximum radiation intensity to the average radiation intensity). There appears to be no significant correlation between the magnitudes of the directivities of these separate secondary sources. The energy combines in a random manner to create the scalloped beam shape.

In making the antenna measurements, the antenna was first boresighted optically on a point midway between the t-r antennas. Then by rotating the pole on which the antenna was mounted, usually by only a fraction of a degree, the exact center position could be found by using a voltmeter to indicate the receiver output. That is to say, since a peak or a null in the beam

pattern, for all cases, occurred right at or only a fraction of a degree away from the optically set position, this corrected position was taken to be the boresight position.

The table of results shows an N (null) or a P (peak) occurring at the boresight position in each data block. For the parabolic, double-dipole feed, X-band antenna polarized the same as the transmitting and receiving horns, nulls occurred whenever the termination was such as to provide good absorption of the waveguide energy. With a termination which did not absorb much of this energy, a peak in the beam pattern occurred. Similarly, nulls occurred in the measured patterns for the X-band and C-band horns when the waveguide was terminated in approximately its characteristic impedance. Generally speaking, peaks in the radiation patterns resulted when the test horns were cross polarized with respect to the t-r horns. The fact that nulls occurred with some X-band antennas cross polarized suggests that the absorption in the termination was not the only factor influencing the radiation characteristics. Interference effects associated with the randomly scattered energy from secondary sources on the antenna can influence the radiation pattern.

Each time a minimum in the beam pattern was observed at the boresight position, a rotation of only three or four degrees in either direction would change the return to a maximum. For the X-band, 18-inch parabolic antenna, these symmetrical peaks were about 25 db above the reference sphere cross section (i.e., on the order of 500 square feet).

Based upon the above observations, it appears that the scattering cross section of an antenna might be expected to change by a factor of 20 or 30 db for a change in orientation of a few degrees with respect to a fixed t-r unit. If the antenna cross section were a detectable portion of the total cross

section of an object on which it was mounted and the antenna were performing some type of mechanical scan, the total scattering cross section could then be expected to fluctuate.

C. Cross Section Amplitudes

Some of the factors which influence the cross section magnitude for the different antennas tested will now be considered.

It must be remembered that these tests were all performed at a transmitter frequency of 9375 Mc/s. This is the center frequency for which the RG-52/U waveguide, used in conjunction with the X-band target antennas, was designed. This frequency is well above the cutoff frequency of the RG-50/U waveguide used for C-band, but below the cutoff of the K-band RG-53/U guide. The cutoff frequency is the frequency at which the width of the guide is a half-wavelength.

Ignoring for the moment just how the energy is supplied, assume that electromagnetic energy in the proper form is made available to the above waveguide transmission lines. Let the attenuation presented this 9375 Mc/s energy in each waveguide now be calculated. For an air filled guide, the attenuation which accompanies propagation may reasonably be taken to result solely from power loss in the guide walls of finite conductivity. Since the waveguide length of interest here is on the order of a foot or so, attenuation per inch will be calculated. For modes having cutoff frequencies less than the signal frequency of 9375 Mc/s, the attenuation due to an imperfect conductor is given by:

$$\alpha_c = \frac{R_s}{b\eta \left[1 - \left(\frac{f_c}{f}\right)^2\right]^{1/2}} \left[1 + \frac{2b}{a} \left(\frac{f_c}{f}\right)^2\right] \quad (7)$$

α_c = nepers/inch

R_s = surface resistivity of the guide material (ohms per square)

b = inside dimension of the narrow side of the guide (inches)

a = inside dimension of the wide side of the guide (inches)

η = intrinsic impedance of the dielectric contained within the guide (ohms)

f_c = cutoff frequency for the mode under consideration

f = signal frequency

The resulting expression may be multiplied by 8.686 to convert it from nepers/inch to db/inch. For both transverse electric (TE_{mn}) and transverse magnetic (TM_{mn}) waves, the cutoff frequency in equation (7) is given by

$$f_c = \frac{1}{2\left[\mu \ \epsilon\right]^{1/2}} \left[\left(\frac{m}{a}\right)^2 + \left(\frac{n}{b}\right)^2\right]^{1/2} \quad (8)$$

μ = permeability of the dielectric

ϵ = dielectric constant or permittivity

m = number of half-sine variations along the "a" direction of the guide

n = number of half-sine variations along the "b" direction of the guide

With a signal frequency of 9375 Mc/s, only three modes have cutoff frequencies below the signal frequency and thus only these three may be propagated along the waveguide. In the C-band guide the TE_{10} and TE_{20} modes could be propagated. In the X-band guide only the TE_{10} mode could be propagated. Propagation cannot occur in the K-band waveguide with this signal frequency.

The values of attenuation in db/inch of waveguide tranversed for a few of the modes having the lowest cutoff frequencies in rectangular guide are given in Table II. The attenuation for modes of higher order than these is progressively greater. Since the waveguide used in each case was at least one foot long, the energy reaching the termination can be neglected for all higher order modes whose attenuations were not calculated.

TABLE II. WAVEGUIDE ATTENUATION (db/inch)

Mode	C-band Guide	X-band Guide	K-band Guide
TE_{10}	2.415×10^{-3}	5.57×10^{-3}	48.2
TE_{01}	6.93	52.75	
TE_{11}	21.1		
TE_{20}	8.1×10^{-3}	42.	
TE_{21}	40.5		
TE_{30}	41.2	79.7	
TM_{11}	21.1	60.5	

If the signal frequency is below cutoff for a particular mode in a particular waveguide, equation (7) cannot be used to calculate the attenuation. Above cutoff where the electromagnetic fields are transmitted, the propagation constant is purely imaginary. Below cutoff this constant is real and a pure attenuation. For the latter condition, again assuming only attenuation due to an imperfect conductor,

$$\alpha_c = \frac{2\pi}{\lambda_c}\left[1 - \left(\frac{f}{f_c}\right)^2\right]^{1/2} \qquad (9)$$

$$\lambda_c = \frac{2ab}{\left[(mb)^2 + (na)^2\right]^{1/2}} \qquad (10)$$

$\pi = 3.1416$

λ_c = cutoff wavelength; the wavelength associated with the cutoff frequency.

These equations were used to calculate the attenuations for all modes other than the TE_{10} and TE_{20} in the C-band guide and the TE_{10} in the X-band guide. Below cutoff, this attenuation is seen from Table II to be of such a magnitude that the field amplitude becomes negligible after only a few wavelengths.

The point of the above discussion is that if the target antenna is properly oriented such that energy in the dominant TE_{10} mode is fed to the waveguide, the C- and X-band guides can be expected to transmit most of this energy to the termination. No energy will be propagated down the K-band guide. Thus the nature of the termination should influence the scattering cross sections of C- and X-band antennas,but not those of the K-band antennas, at this signal frequency.

To say that electromagnetic waves are attenuated by so many db/inch is not sufficient here, since the chief concern is not how much energy gets to the termination, but rather what portion of this energy is re-radiated towards the receiver. In other words, a physical interpretation must be given to the statement that energy is attenuated. Is the signal absorbed at that point at which it meets a waveguide for which it is below cutoff or is the energy reflected from this point? The fact that energy does not progress beyond this point is not questioned.

An extension of the theory of propagation down guides of constant cross section to apply to tapered waveguides suggests a solution which agrees neither with the empirical results nor with the anticipated results.

The velocity with which the electric and magnetic fields advance down a waveguide is known as the group velocity.* This velocity is related to v, the free space velocity of electromagnetic energy, by the relationship

$$V_g = V \sin \theta$$

where θ is the angle which the plane wavefront of the electromagnetic fields makes with the normal to the side of the guide. For θ equal to zero, the wave would be propagating directly into the walls of the guide and the corresponding group velocity would be zero. This is the condition which exists as the signal frequency reaches the cutoff frequency of the waveguide. Theoretically, as the signal is decreased toward cutoff (the wavelength is increased) the angle θ becomes progressively smaller. This means that for a given axial distance transversed by the wave, the number of times that the energy bounces back and forth from one side of the guide to the other is increased as the frequency is decreased. Visualized in this manner, in the limit (at cutoff) the electromagnetic energy would be bouncing normally to the walls and would not be progressing axially.

Since the conductivity of the waveguide walls is finite, a portion of the energy of the electromagnetic fields will be absorbed. If, as hypothesized above, the fields advance to a certain plane beyond which they cannot advance and in which the energy rings back and forth, then eventually all of the energy of the wave should be dissipated as heat in the guide walls. To assume that this is the situation, then, is to assume that a waveguide below cutoff presents a matched termination from which, of course, no energy will be reflected. This conclusion is not in agreement with the empirical results.

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*Jordan, E. C., Electromagnetic Waves and Radiating Systems, Prentice-Hall Inc., N. Y., 1950, p. 689.

The applicability of equation (11) to horns or tapered waveguide is the point in question. Referring to the data in Table I, the X- and K-band horns appear to present an almost identical scattering cross section. The horns were of the same physical dimensions, the only difference being that they were tapered so as to terminate in different sized waveguides. The point of significance is that the K-band horn appears the same as an X-band horn shorted, not properly terminated. On the assumption that most of the energy which enters the shorted X-band horn comes back out, the data would thus suggest that at the plane at which the K-band horn tapers to below cutoff the impedance presented is more like a short than a matched termination.

Although an analytical proof will not be offered here, some of the reasoning will be given which supports the idea that a horn which tapers to below cutoff will appear to be shorted rather than properly terminated.

Continuing with the ray diagram representation of the electromagnetic energy, consider energy incident upon a horn from free space. In matching the free space fields to the waveguide, the free space wavelength must be increased to the guide wavelength. That this is suggested in Fig. 12, where with each bounce the angle between the ray and the normal to the side of the horn is decreased. The axial advancement of the wave between bounces is progressively reduced, much the same as it is reduced in a guide of constant dimensions as the wavelength is increased (i.e., as cutoff is approached).

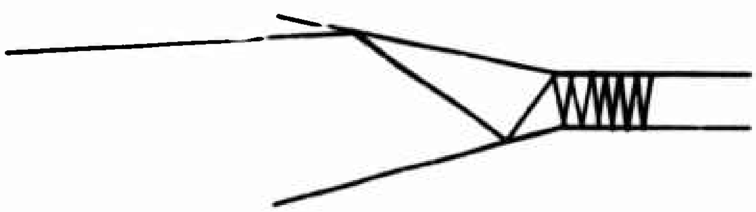

Fig. 12. Ray Diagram Representation of Waves in a Horn

Figure 13 shows in more detail the angles which the ray would make with the sides of the horn. This angle is seen to increase with each bounce. If energy were incident at an angle ϕ from the axial direction of the horn, then, considering ray number one of this illustration, in order to couple energy into the waveguide the guide would have to be tied to the horn at section B-B or sooner. Otherwise the energy would pass through some angle such that on the next bounce it would be directed back out of the horn. Considering ray number two, on the other hand, it would appear that energy incident nearer the throat of the horn could be successfully coupled into a guide at section A-A. Obviously this guide would be much smaller than the guide at section B-B.

As the waveguide used to couple the energy out of the horn is reduced still further in size, it becomes apparent that the ability of the horn to funnel energy into the guide is practically nil. In other words it ceases to act as an effective antenna. The only energy which gets into the guide is that which is incident directly upon the guide itself. Here the cutoff criterion discussed earlier for waveguide of constant cross section does apply.

The fact that the angle of the energy ray must change with each bounce in a horn or tapered waveguide suggests that the energy could not reach some plane and remain there, ringing back and forth between the walls, until it had all been dissipated as heat. The angles of the horn force the energy to have an axial propagation component.

As the number of bounces per unit length of advancement increases as the wave moves toward a guide of reduced cross sectional area, so does the energy loss per unit length. For a horn of the size considered in the experimental part of this study, where the number of bounces involved before

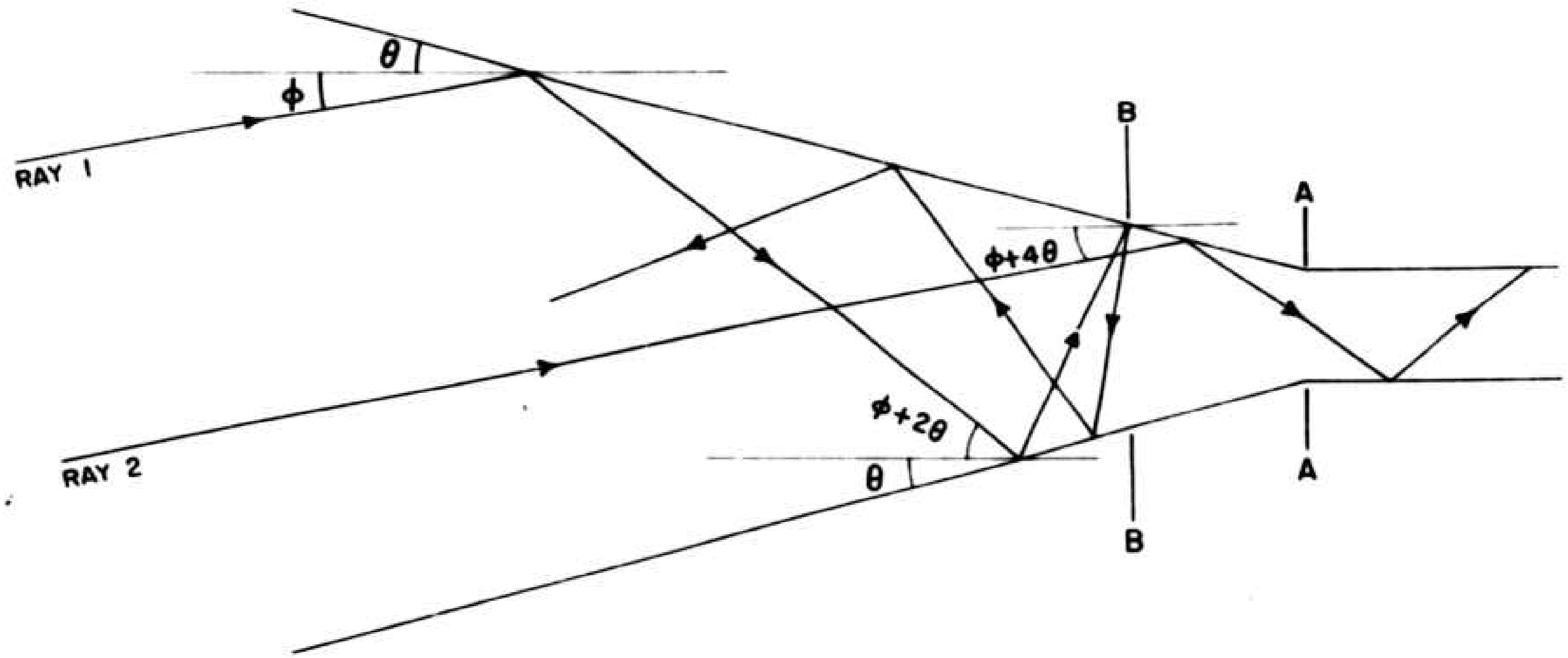

FIGURE 13

EXPANDED RAY DIAGRAM REPRESENTATION IN A HORN

the energy is either reflected back out or coupled into a waveguide is relatively small, one would expect the ratio of energy absorbed to energy reflected to be small for signals below cutoff. Thinking again of the idea of scattering cross section, a horn coupled to a waveguide below cutoff should thus appear terminated in approximately a short circuit.

For a tapered waveguide, if the transition from a size above cutoff to one below cutoff were extended over a long distance, then the energy would pass an extended region over which conductor losses would be high. Such a device, tapering to zero cross section, might thus be used to provide a matched termination for all modes and all frequencies.

In addition to the absorption and reflection properties of the waveguides and terminations themselves, other factors influence the scattering cross sections of the parabolic reflectors.

A two dipole feed system was used for each of the parabolic reflectors tested. The feed system, as seen in Fig. 1, consisted of a resonant dipole and a parasitically excited dipole, both mounted on a web in the center of the waveguide. The resonant or driven dipole, the one nearest the waveguide, was located at the focal point of the parabolic reflector. This focal point location should be insensitive to signal frequency variations anywhere in the radar band. This type of feed is frequency sensitive, however. The feed is a simplified Yagi antenna having only a driven element and a reflector. The radiation pattern of the combination is determined by the spacing, in wavelengths, between the dipoles as well as by the length and diameter of the dipoles.

The feeds of the parabolic antennas tested were designed to handle different wavelengths, but were all illuminated with energy having a 3.2 centimeter wavelength. The dipoles were spaced 0.4 λ apart for the X-band

feed, 0.595 λ for the C-band feed, and 0.169 λ for K-band feed. In terms of signal wavelength the length of the driven dipole was 0.46 λ for X-band, 0.59 λ for C-band, and 0.194 λ for K-band.

The net effect of these variations in spacing and length was to increase the amount of energy which was radiated by the dipole in a direction other than back to the waveguide or back to the parabolic reflector. Thus the C- and K-band antenna feeds were not as efficient as the X-band feed in coupling into the waveguide that energy which was focused upon the driven dipole.

Since more energy got into the X-band guide, if the guide were shorted or otherwise provided a poor impedance match, more energy should have come back out of the X-band guide, been focused on the parabolic reflector, and been directed back to the t-r mount. Since the X-band feed couples most of the incident X-band signal into the waveguide, the influence on the cross section of this focused energy should far overshadow that which never entered the guide. Thus, with a poorly terminated waveguide, the X-band parabolic reflector should present a much larger cross section than either the C-band or the K-band antenna. On the other hand, if the guide were properly terminated, most of the X-band energy would feed into the waveguide, would be absorbed, and would not return. A smaller percentage of the total would get into the C-band guide and none would get into the K-band guide, so less would be absorbed and more would be available to be scattered back to the receiver. Thus, when properly terminated, the X-band antenna might be expected to have a smaller scattering cross section than either the C-band or the K-band antenna.

Thus far, the fact that the cross sections given in Table I are, at least, of the proper order of magnitude has not been established analytically. An estimate of the cross section magnitude of a shorted 18-inch parabolic reflector will now be made. Only the shorted termination will be considered because it may then be assumed that, for all practical purposes, all of the energy getting into the waveguide will be reflected back out. For other terminations, it is more difficult to establish just what percentage of the energy getting in to the guide will be returned.

The effective receiving cross section or effective aperture of this antenna is

$$A_e = f\,A = 1.06 \text{ sq. ft.} \tag{12}$$

$f = 0.6$; a factor indicating the uniformity of the phase and intensity over the parabolic aperture

$A = \frac{\pi D^2}{4} = 1.76$ sq. ft.; the area of the parabolic aperture

The power gain of this antenna above an isotropic source is

$$G = \frac{4\pi A_e}{\lambda^2} = 1208 = 30.8 \text{ db} \tag{13}$$

G = power gain relative to an isotropic radiator

$\lambda = 3.2$ cm, the signal free space wavelength

The effective aperture is associated with the energy of the incident signal which gets to the waveguide feed. For an X-band antenna and an X-band signal, a large portion of the energy getting to the feed is coupled to the antenna termination. The scattering cross section is determined by energy which is re-radiated from the antenna. Thus the power gain of the antenna

should be the factor which relates the effective aperture to the scattering cross section, A_s,

$$A_s = G\ A_e \tag{14}$$

$$A_s = 1208 \text{ sq. ft.}$$

The empirically established value for this antenna and termination was about 5.9 db above this value. However, it must be emphasized that a number of simplifying assumptions were made in this calculation. First, energy which was scattered from the antenna without getting to the feed system was not considered. Second, losses at the feed were neglected. Third, the error in the empirical value caused by ringing, as discussed earlier, was not considered. None the less, the empirical value does seem a little large.

V. SUMMARY AND CONCLUSIONS

The radar technique employed in performing the antenna scattering cross section tests, that of cancelling the returns from fixed background objects, was successful. The amount of microwave absorbent material required for the test was considerably less than that which would have been required in building an anechoic chamber.

It was not possible in the tests to measure with high accuracy the radiation patterns or scattering cross sections of the antennas when pointed in any direction other than directly at the transmitter-receiver unit. However, in this boresight position the results appear credible. The effect of multiple bouncing of the transmitted energy between the test antenna and the transmitter-receiver unit is of little consequence, even for a target range of only twenty feet as used in this experiment.

The radiation pattern of an antenna acting as a passive scatterer has a scalloped appearance and contains more inflections than would be found for the free space beam pattern of the same antenna functioning as a transmitter or a receiver. Apparently the antenna scattering cross section could change by a factor of 20 or 30 db for a change in orientation with respect to a fixed transmitter-receiver unit of just a few degrees.

When antennas designed for C-, X-, and K-band are illuminated with X-band energy, the influence of the terminating impedance on the scattering cross section is more significant for the X-band antenna. If the transmission line associated with the antenna is mismatched, an X-band parabolic antenna, differing from a C- or K-band antenna only in the feed design, should give the largest return. For the waveguide terminated in a matched impedance, the X-band scattering cross section will be the smaller of the three.

The terminating impedance of a K-band antenna will not influence the scattering cross section if the frequency of the illuminating signal is in X-band. For such a situation, a K-band horn of the size tested in this experiment will always appear to be terminated in a short circuit.

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